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SELF-ADJUSTING MODULATOR

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(71) Applicant(s)
ERICSSON INC.

(72) Inventor(s)
PAUL W. DENT

(74) Attorney or Agent
WATERMARK PATENT & TRADEMARK ATTORNEYS , Locked Bag 5, HAWTHORN VIC 3122

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(57) Claim

1.   A self-adjusting modulator for providing an accurately modulated signal, comprising:

first digital signal processing means for numerically generating in-phase and quadrature modulation waveform samples and applying correction factors to numerically generated in-phase and quadrature modulation waveform samples;

a quadrature modulator for impressing an in-phase modulation waveform, corresponding to the numerically generated in-phase modulation waveform samples, on a substantially cosinusoidal carrier wave and for impressing a quadrature modulation waveform, corresponding to the numerically generated quadrature modulation waveform samples, on a substantially sinusoidal carrier wave;

transmitter means for upconverting and amplifying an output of said quadrature modulator to a predetermined transmission frequency and power level;

modulation assessment receiving means, coupled to said transmitter means, for collecting and digitising samples of the upconverted and amplified output of said quadrature modulator; and

second digital signal processing means, in communication with said first digital signal processing means and with said modulation assessment receiving

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means, for determining the correction factors based on said digitised samples from said modulation assessment receiving means and on uncorrected in-phase and quadrature modulation waveform samples from said first digital signal processing means, and for communicating the correction factors to the first digital signal processing means.

13.   A self-adjusting modulation method for providing an accurately modulated signal, comprising the steps of:

a)   numerically generating uncorrected in-phase and quadrature modulation waveform samples;

b)   applying a set of predetermined correction factors to said uncorrected in-phase and quadrature modulation waveform samples to produce corrected in-phase and quadrature modulation samples;

c)   forming a corrected in-phase modulation waveform corresponding to the corrected in-phase modulation samples and modulating a substantially cosinusoidal carrier wave with the corrected in-phase modulation waveform, forming a corrected quadrature modulation waveform corresponding to the corrected quadrature modulation samples and modulating a substantially sinusoidal carrier wave with the corrected quadrature modulation waveform, and forming a modulated output signal by summing the modulated cosinusoidal and sinusoidal carrier waves;

d)   upconverting and amplifying the modulated output signal to a selected desired transmission frequency and power level;

e)   collecting digital samples of the upconverted, amplified modulated output signal;

f)   processing said digital samples with the uncorrected in-phase and quadrature waveform samples to produce an updated set of correction factors and

g)   successively repeating the foregoing steps a), b), c), and d).

(54) Title: SELF-ADJUSTING MODULATOR

(57) Abstract

The method and apparatus involve receiving (118) one's own transmission with a suitable modulation assessment receiver (119-125) and determining (125) the modulation error relative to the theoretically perfect modulation transmission expected by the receiver. The measured error is used to adjust (110) the modulation to minimize the error. The modulation assessment receiver may use log-polar signal processing (122) to measure phase (124) and (log)amplitude (123) instead of cartesian I and Q components, and then convert (125) to cartesian form.

SELF-ADJUSTING MODULATOR

The present invention relates to radio transmitters optimized for digital data transmission, and more particularly, to improving the accuracy with which digital data may be impressed on a radio carrier frequency wave by means of a quadrature modulator. Improved accuracy of digital data impression on radio carrier frequency waves is especially important and useful for recent developments in Viterbi, echo-integrating demodulators and in subtractive demodulation of Code Division Multiple Access (CDMA) modulations.

BACKGROUND

The cellular telephone industry has made phenomenal strides in commercial operations in the United States as well as the rest of the world. Growth in major metropolitan areas has far exceeded expectations and is outstripping system capacity. If this trend continues, the effects of rapid growth will soon reach even the smallest markets. Innovative solutions are required to meet these increasing capacity needs as well as maintain high quality service and avoid rising prices.

Throughout the world, one important step in cellular systems is to change from analog to digital transmission. Equally important is the choice of an effective digital transmission scheme for implementing the next generation of cellular technology. Furthermore, it is widely believed that the first generation of Personal Communication Networks (PCNs), (employing low cost, pocket-size, cordless telephones that can be carried comfortably and used to make or receive calls in the home, office, street, car, etc.), would be provided by the cellular carriers using the next generation digital cellular system infrastructure and the cellular frequencies. The key feature demanded in these new systems is increased traffic capacity.

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Currently, channel access is achieved using Frequency Division
Multiple Access (FDMA) and Time Division Multiple Access (TDMA) methods.
As illustrated in Fig. 1(a), in FDMA, a communication channel is a single radio
frequency band into which a signal's transmission power is concentrated.
5   Interference with adjacent channels is limited by the use of band pass filters which
only pass signal energy within the specified frequency band. Thus, with each
channel being assigned a different frequency, system capacity is limited by the
available frequencies as well as by limitations imposed by channel reuse.

In TDMA systems, as shown in Fig. 1(b), a channel consists of a
10   time slot in a periodic train of time intervals over the same frequency. Each
period of time slots is called a frame. A given signal's energy is confined to one
of these time slots. Adjacent channel interference is limited by the use of a time
gate or other synchronization element that only passes signal energy received at
the proper time. Thus, the problem of interference from different relative signal
15   strength levels is reduced.

Capacity in a TDMA system is increased by compressing the
transmission signal into a shorter time slot. As a result, the information must be
transmitted at a correspondingly faster burst rate which increases the amount of
occupied spectrum proportionally. The frequency bandwidths occupied are thus
20   broader in Fig. 1(b) than in Fig. 1(a).

With FDMA or TDMA systems or hybrid FDMA/TDMA systems,
the goal is to ensure that two potentially interfering signals do not occupy the same
frequency at the same time. In contrast, Code Division Multiple Access (CDMA)
allows signals to overlap in both time and frequency, as illustrated in Fig. 1(c).
25   Thus, all CDMA signals share the same frequency spectrum. In either the
frequency or the time domain, the multiple access signals overlap. In principle,
the informational datastream to be transmitted is impressed upon a much higher bit
rate datastream generated by a pseudorandom code generator. The informational
datastream and the high bit rate datastream are multiplied together. This
30   combination of higher bit rate signal with the lower bit rate datastream is called
coding or spreading the informational datastream signal. Each informational
datastream or channel is allocated a unique spreading code. A plurality of coded

information signals are transmitted on radio frequency carrier waves and jointly received as a composite signal at a receiver. Each of the coded signals overlaps all of the other coded signals, as well as noise-related signals, in both frequency and time. By correlating the composite signal with one of the unique spreading codes, the corresponding information signal is isolated and decoded.

There are a number of advantages associated with CDMA communication techniques. The capacity limits of CDMA-based cellular systems are projected to be up to twenty times that of existing analog technology as a result of the properties of a wide band CDMA system, such as improved coding gain/modulation density, voice activity gating, sectorization and reuse of the same spectrum in every cell. CDMA is virtually immune to multi-path interference, and eliminates fading and static to enhance performance in urban areas. CDMA transmission of voice by a high bit rate decoder ensures superior, realistic voice quality. CDMA also provides for variable data rates allowing many different grades of voice quality to be offered. The scrambled signal format of CDMA completely eliminates cross talk and makes it very difficult and costly to eavesdrop or track calls, insuring greater privacy for callers and greater immunity from air time fraud. Various aspects of CDMA communications are described in K. Gilhousen et al., "On the Capacity of a Cellular CDMA System", IEEE Trans. on Vehicular Technology vol. 40, pp. 303-312 (May 1991).

In systems optimized for digital data transmission, M-ary digital modulation methods, in which one of M possible signals are transmitted during each signalling interval, are often used because of their improved efficiency. One method commonly used is Quadrature Phase Shift Keying (QPSK) in which two equal-magnitude signals in phase-quadrature are impressed on the carrier wave. Another common method is Offset QPSK (OQPSK) in which the maximum phase transition at any point in the modulated waveform is less than the maximum phase transition in a QPSK waveform. As a result, the composite OQPSK signal can have smaller envelope fluctuations after bandpass filtering, and thus more closely approach the constant envelope desired for such signalling. It will be appreciated that QPSK and OQPSK are forms of Quadrature Amplitude Modulation (QAM). Various aspects of these modulations are set forth in, for example, F. Stremler,

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Introduction to Communication Systems, 2d ed., pp. 590-596, Addison-Wesley Publishing Co., Reading, Massachusetts (1982); and S. Gronemeyer et al., "MSK and Offset QPSK Modulation", *IEEE Trans. on Communications* vol. COM-24, pp. 809-820 (Aug. 1976).

5   An example of a system optimized for digital data transmission is a CDMA system in which the demodulation of quadrature-modulated signals involves comparing the received wave with a theoretical wave modulated with hypothesized data patterns, e.g., a Viterbi demodulator. Another example of such a system is a CDMA system in which a stronger signal is demodulated first and

10   then is subtracted from the received signal before a remaining weaker signal is demodulated, as described in commonly assigned U.S. Patent Nos. 5,151,919 and 5,218,619. Both of these documents are expressly incorporated here by reference.

             A typical quadrature modulator takes advantage of the quadrature phases of sine and cosine waves to modulate twice the information rate on the

15   radio carrier wave. For example, the even bits in a digital information datastream can be modulated on the cosine wave, and the odd bits in the digital information datastream can be modulated on the sine wave. Errors can arise in quadrature modulators whenever the phases of the cosine and sine waves are not quite 90° apart, or whenever the amplitudes of the sine and cosine waves are not quite

20   equal, or whenever there is residual carrier leakage when the modulating wave is supposedly zero, as well as for other reasons. Of course, the accuracy with which the quadrature modulation matches a synthesized, theoretical wave modulated with hypothesized data or with already received data is important in the above-described communication systems. The accuracy of quadrature modulators has been

25   maintained conventionally by a combination of ensuring good matching between components and by making trimming adjustments to reduce residual mismatch errors.

             A conventional quadrature modulator, shown in Fig. 2, includes an "in-phase" or I modulator 101, a "quadrature" or Q modulator 102, and a phase-

30   splitting network 103 for supplying the double-sideband, suppressed carrier modulators 101, 102 with cosine and sine carrier frequency signals, respectively. Ideally, the signals provided by the network 103 are cos(ωt) and sin(ωt), where ω

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is the carrier signal's angular frequency. Also shown in Fig. 2 are an I and Q
modulation generator 104 for supplying I and Q modulation signals, a combination
network 105 for adding the outputs of the I modulator 101 and the Q modulator
102, and trim potentiometers 106, 107 for carrier balance/d.c. offset adjustments

5   for the I and Q signals, respectively. Additional trim potentiometers 108, 109 for
amplitude matching the I and Q signals, respectively, are also shown in Fig. 2.
The phase-splitting network 103 may also be adjustable, as indicated by the
diagonal arrow, to achieve as nearly as possible the desired 90° phase difference
between the sine and cosine carrier frequency signals.

10   In practice, if the I modulator 101 and Q modulator 102 are
constructed on the same silicon chip by integrated circuit technology, they will be
very well matched, possibly eliminating the need for the amplitude adjustment
potentiometers 108, 109. Also in some cases, the purposes of the phase-splitting
network 103 can be achieved by starting with a signal having a frequency of 4ω,

15   i.e., four times the desired carrier frequency ω, and using the 4ω-signal to clock a
digital logic divide-by-four circuit that produces the bit patterns:

<p style="text-align:center">0011001100110011...</p>

and

<p style="text-align:center">0110011001100110...,</p>

20   which can be recognized as square waves of frequency 1/4 the 4ω bitrate having
phases exactly 1/4 of a period (90°) apart. It is usually acceptable to drive the I
and Q modulators with square-wave carrier signals instead of with sinusoidal
signals. Such a digital method of producing 90°-phased signals can be practical
for frequencies ranging up to hundreds of megahertz, but at higher frequencies

25   slight differences in speed of loading of the logic circuits can once more become a
significant source of modulator error.

            The carrier balance and/or d.c. offset adjustments try to ensure that,
when the modulation generator 104 produces a zero signal level on its I and Q
outputs, the corresponding output at the carrier frequency of the I and Q

30   modulators is also zero. In essence, this requires the I modulator 101 to produce
a zero cosine signal output for a zero I modulation and the Q modulator to
produce a zero sine signal output for a zero Q modulation. It is well known that

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an I modulator imbalance can actually produce a sine signal when the cosine signal
is zero, and a Q modulator imbalance can produce a cosine signal when the sine
signal is zero. Accordingly, a small cosine leakage from the I modulator is
sometimes desired to balance a cosine leakage from the Q modulator, and a small
5   sine leakage from the Q modulator is sometimes desired to balance a sine leakage
from the I modulator. With the two adjustment potentiometers 106, 107,
however, carrier balance is more readily achievable.

Other sources of modulation inaccuracies are non-linearity in the
modulators 101, 102 and non-linearity in the modulation generator 104. The
10   generator 104 often produces precursors of the I and Q modulation signals
numerically by means of a digital signal processor, and then converts the
precursors to analog modulating signals be means of digital-to-analog (D/A)
converters. Mismatches between the I-signal D/A converter and the Q-signal D/A
converter or in the anti-aliasing filters thereafter are a further source of modulation
15   error. In some cases, the digital signal processor computes a pre-distortion of the
modulating signal using an inverse of the non-linear transfer functions of the
modulators 101, 102 in order to compensate for modulator non-linearity.
Techniques for simplifying the D/A conversion and subsequent anti-aliasing
filtering by use of highly oversampled delta-modulation are also known, and lead
20   to some reduction of the aforementioned modulation errors. One such technique is
described in commonly assigned U.S. Patent Application No. 07/967,027 entitled
"Multi-Mode Signal Processing", which is expressly incorporated here by
reference.

U.S. Patent No. 4,985,688 to Nagata discloses a modulation system
25   in which an amplified, modulated output signal is fed back to a quadrature
demodulator. The signal is demodulated and compared to a threshold value. A
control signal is generated based on this comparison to adjust the system for
nonlinearities of the amplifier that is connected to the modulator. When the
threshold is exceeded, the normal modulation is apparently interrupted and
30   replaced by a signal of 1/N-th the frequency or data rate. The Nagata patent also
describes how to determine the instants at which the output of the quadrature

demodulator should be sampled by use of a differentiator, divider circuit, and clock controlling means.

The Nagata patent's device may also be described as an adaptive, self-learning predistortion arrangement. The stated purpose of the Nagata patent is to inverse-predistort the input to a quadrature modulator such that the output after a distorting power amplifier is correct. On the other hand, the Nagata patent's device can hardly correct for errors in a quadrature modulator because it uses a quadrature demodulator to assess errors, and as described above the demodulator is likely to suffer from the same type of errors as the modulator. After all, if one could make a perfect demodulator, one would simply use it as a perfect modulator.

U.S. Patent No. 4,581,749 to Carney et al. discloses a frequency modulation device usable in a mobile communication system. A feedback loop provides control of angle modulation error by comparing the modulated deviation amount with a predetermined deviation value. The automatic modulation error correction system described is for transmitters using pure angle modulation, specifically binary continuous-phase frequency shift keying (CPFSK).

In the system described in the Carney patent, an exact modulation index is generated by digitally switching the frequency between two exact values. Nevertheless, such a modulation is not used for transmission because the transitions are not filtered. The transmit waveform uses shaped one-zero transitions to contain the spectrum, and when a sufficient number of like bits occur in a row the frequency deviation of the shaped modulation should approach the same value as the unshaped modulation. Occurrences of such strings of like bits are detected and a comparison is made when they occur, the result being used in a feedback loop to adjust the modulation index. Thus, the Carney patent assesses errors only when the modulation is a long-enough string of ones or zeroes.

U.S. Patent No. 5,020,076 to Cahill et al. describes switching between analog FM modulating a carrier signal source in the conventional way, and modulating it using a quadrature modulator. The quadrature modulator is left in the circuit when conventional FM is carried out, and the I and/or Q modulation signal is just set to a constant so that the quadrature modulator passes the FM signal straight through.

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U.S. Patent No. 4,856,025 to Takai describes a transmit diversity implementation for improving digital radio communication. A special waveform and special receiver are used, but the special receiver does not assess the accuracy of the transmitter modulation to provide information to a modulation correction

5   system.

It will be appreciated from the foregoing that high modulation accuracy has heretofore been achieved by good design practices combined with specific, fixed, once-and-for-all adjustments that can compensate for fixed, immutable imperfections. It would be highly desirable to be able, continuously

10   and interactively, to adjust and compensate for mutable modulation inaccuracies and errors as well.

SUMMARY

In accordance with the present invention, a transmitter receives its own transmission with an appropriate receiver and determines the transmission's

15   modulation errors relative to the theoretical form the receiver expects. Adjustments are made interactively to the modulating waves in a direction that reduces the errors until convergence to the desired, theoretical form is achieved.

In one embodiment, a digital signal processor generates the I and Q modulation waveforms numerically, and makes numerical adjustments by adding

20   offsets to achieve carrier balance, by multiplicative scaling to achieve I and Q matching, and by I and Q cross-coupling to compensate for $90°$-phase-splitting errors. The numerical adjustments are continuously updated by a modulation assessment receiver operating on a sample of the transmitter output.

If a particular type of modulator (or demodulator) is susceptible to

25   particular types of modulation inaccuracy, that modulator can be adapted to measure and report the features of the modulation necessary to continuously update the numerical adjustments, particularly when presented with a noise-free sample of the transmitted signal to evaluate. A modulation assessment receiver is provided for a system in which the modulation is a spread-spectrum signal using orthogonal

30   or bi-orthogonal coding.

BRIEF DESCRIPTION OF THE DRAWINGS

The features and advantages of the invention will be understood by reading the following detailed description in conjunction with the drawings, in which:

5   Figs. 1(a)-(c) are plots of access channels using different multiple access techniques;

Fig. 2 is a functional block diagram of a typical quadrature modulator;

Fig. 3 is a functional block diagram of a system in accordance with 10   the present invention;

Fig. 4 illustrates how CDMA signals are generated;

Figs. 5 and 6 illustrate how CDMA signals are decoded;

Fig. 7 illustrates a subtractive CDMA demodulation technique;

Figs. 8(a), 8(b) are block diagrams of a transmitter and a receiver in 15   a spread-spectrum communications system; and

Fig. 9 shows waveforms of Shaped Offset Quadrature Amplitude Modulation (SOQAM).

DETAILED DESCRIPTION

While the following description is in the context of cellular 20   communications systems involving portable or mobile radio telephones and/or Personal Communication Networks (PCNs), it will be understood by those skilled in the art that the present invention may be applied to other communications applications. Moreover, while the present invention may be used in a subtractive CDMA demodulation system, it also may be used in applications of other types of 25   spread-spectrum systems.

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QUADRATURE MODULATOR IMPERFECTIONS

To assist in understanding the present invention, an analysis of the imperfections that can arise in quadrature modulators will first be given.

If the I and Q channels have a gain imbalance such that the gain of one channel is a factor A higher than the geometric mean, $(IQ)^{1/2}$, of the amplitude I of the unmodulated I-channel input and the amplitude Q of the unmodulated Q-channel input) and gain of the other channel is a factor A lower than $(IQ)^{1/2}$, and if, in addition, the cosine and sine carrier signals are not exactly 90° apart but a phase error $+\delta\phi$ exists on the one carrier signal and a phase error $-\delta\phi$ exists on the other carrier signal relative to some mean phase, then the cartesian form of the modulator output signal can be written as:

$$AI\cos(\omega t + \delta\phi) + (Q/A)\sin(\omega t - \delta\phi),$$

which expands to:

$$\cos(\omega t)[\cos(\delta\phi)AI - (\sin(\delta\phi)/A)Q] + \sin(\omega t)[(\cos(\delta\phi)/A)Q - \sin(\delta\phi)AI].$$

Defining $C = \cos(\delta\phi)$ and $S = \sin(\delta\phi)$, and using matrix notation, the modulator output signal becomes:

$$[\cos(\omega t) \quad \sin(\omega t)] \begin{bmatrix} C & -S \\ -S & C \end{bmatrix} \begin{bmatrix} A & 0 \\ 0 & 1/A \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix}.$$

Since a desirable form is:

$$[\cos(\omega t) \quad \sin(\omega t)] \begin{bmatrix} I_0 \\ Q_0 \end{bmatrix},$$

it can be seen that the following is required:

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$$\begin{bmatrix} C & -S \\ -S & C \end{bmatrix} \begin{bmatrix} A & 0 \\ 0 & 1/A \end{bmatrix} \begin{bmatrix} I \\ Q \end{bmatrix} = \begin{bmatrix} I_0 \\ Q_0 \end{bmatrix},$$

which is satisfied when:

$$\begin{bmatrix} I \\ Q \end{bmatrix} = \frac{1}{C^2 - S^2} \begin{bmatrix} 1/A & 0 \\ 0 & A \end{bmatrix} \begin{bmatrix} C & S \\ S & C \end{bmatrix} \begin{bmatrix} I_0 \\ Q_0 \end{bmatrix}.$$

Since the common factor $\lambda (C^2 - S^2)$ is just an overall amplitude scaling, it can be dropped. Similarly, the right-hand side can be divided by $C = \cos(\delta\phi)$, and $T = \tan(\delta\phi) = S/C$ can be defined, with the following result:

$$\begin{bmatrix} I \\ Q \end{bmatrix} = \begin{bmatrix} 1/A & 0 \\ 0 & A \end{bmatrix} \begin{bmatrix} 1 & T \\ T & 1 \end{bmatrix} \begin{bmatrix} I_0 \\ Q_0 \end{bmatrix},$$

5   which can be written in non-matrix form as:

$$I = (I_0 + TQ_0)/A$$

and

$$Q = (Q_0 + TI_0)A.$$

10   The foregoing mathematical analysis demonstrates that the desired modulation may be obtained in spite of modulator imperfections by adding a fraction T of Q_0 to I_0 and then dividing by A and by adding a fraction T of I_0 to Q_0 and then multiplying by A.

One imperfection omitted from the foregoing analysis is a carrier imbalance, which can be taken into account simply by subtracting from I and Q

15   the constants K_i and K_q equal to the amount of carrier imbalance to be corrected. The final formulae for the cartesian form of the I and Q are thus:

$$I = (I_0 + TQ_0)/A - K_i$$

and

$$Q = (Q_0 + TI_0)A - K_q.$$

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In accordance with the present invention, the correction factors A, T, K_i, and K_q are determined by sampling the modulator output waveform with a modulation assessment receiver and communicating the samples to a digital signal processor for generating the I and Q modulations and performing the

5   precorrections given above.

Generally, the modulation assessment receiver must have a means for measuring the I and Q values actually generated by the transmitter's quadrature modulator and a means for comparing the measured I and Q values with ideal I and Q values in order to determine the correction factors. A conventional

10   receiver, however, normally resolves a radio signal into I and Q components using the same type of quadrature modulator circuit that the transmitter uses but operating in reverse. As described above, it is impossible, in principle, to distinguish errors in the modulator from errors in the demodulator in this case.

To avoid this error indistinguishability, a modulation assessment

15   receiver in accordance with one aspect of the present invention uses log-polar signal processing to measure the transmitter signal's phase and the logarithm of its amplitude, instead of the cartesian I and Q components. After digitization, the receiver numerically converts the measurements from the log-polar form to the desired cartesian form. Log-polar signal processing is described in U.S. Patent

20   No. 5,048,059, which is expressly incorporated here by reference.

A simple procedure for estimating the correction factors, given a set of desired I and Q values, I_j and Q_j, for $j = 1, \ldots, N$, and a set of corresponding measured I and Q values, I'_j and Q'_j, is as follows, where it is postulated that:

25                               $$I'_j = BI_j - BTQ_j + K_i \tag{1}$$

and

                               $$Q'_j = CQ_j - CTI_j + K_q, \tag{2}$$

and it is desired to find B, C, $T = \tan(\delta\phi)$, K_i, and K_q that satisfy Eqs. (1) and (2).

First, the desired I_j and Q_j are divided into two subsets, the first

30   subset $\{I_j^{pos}, Q_j^{pos}\}$ containing only the positive values of I_j, and the second subset

$\{I_j^{neg I}, Q_j^{neg I}\}$ containing only the negative values of I_j. If one subset contains more values than the other subset, only N_1 values (where N_1 is equal to the number of values in the smaller subset) are used in both subsets.

The sum over N_1 values of I_j in the positive subset $\{I_j^{pos I}, Q_j^{pos I}\}$ is defined to be I_{S1}:

$$I_{S1} = \sum_{j=1}^{N_1} I_j^{pos I} \ ,$$

and the sum over N_1 values of Q_j in the positive subset $\{I_j^{pos I}, Q_j^{pos I}\}$ is defined to be Q_{S1}:

$$Q_{S1} = \sum_{j=1}^{N_1} Q_j^{pos I}.$$

The sum over N_1 values of I'_j in the subset $\{I'_j^{pos I}, Q'_j^{pos I}\}$ corresponding to the positive subset $\{I_j^{pos I}, Q_j^{pos I}\}$ is defined to be I'_{S1}:

$$I'_{S1} = \sum_{j=1}^{N_1} I'_j^{pos I} \ ,$$

and the sum over N_1 values of Q'_j in the subset $\{I'_j^{pos I}, Q'_j^{pos I}\}$ corresponding to the positive subset $\{I_j^{pos I}, Q_j^{pos I}\}$ is defined to be Q'_{S1}:

$$Q'_{S1} = \sum_{j=1}^{N_1} Q'_j^{pos I} \ .$$

Because all of the I values summed to get I_{S1} were chosen to be positive and the signs of the corresponding Q values summed to get Q_{S1} were uncorrelated, I_{S1} will in general be much larger than Q_{S1}.

The sum over N_1 values of I_j in the negative subset $\{I_j^{neg I}, Q_j^{neg I}\}$ is defined to be I_{S2}:

$$I_{S2} = \sum_{j=1}^{N_1} I_j^{neg I}$$

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and the sum over N_1 values of Q_j in the negative subset $\{I_j^{neg I}, Q_j^{neg I}\}$ is defined to be Q_{S2}:

$$Q_{S2} = \sum_{j=1}^{N_1} Q_j^{neg I}.$$

In a similar way, the sum over N_1 values of I'_j in the subset $\{I'_j^{neg I}, Q'_j^{neg I}\}$ corresponding to the negative subset $\{I_j^{neg I}, Q_j^{neg I}\}$ is defined to be I'_{S2}:

$$I'_{S2} = \sum_{j=1}^{N_1} I'_j^{neg I}$$

5   and the sum over N_1 values of Q'_j in the subset $\{I'_j^{neg I}, Q'_j^{neg I}\}$ corresponding to the negative subset $\{I_j^{neg I}, Q_j^{neg I}\}$ is defined to be Q'_{S2}:

$$Q'_{S2} = \sum_{j=1}^{N_1} Q'_j^{neg I}.$$

Because all of the I values summed to get I_{S2} were chosen to be negative and the signs of the corresponding Q values summed to get Q_{S2} were uncorrelated, the magnitude of I_{S2} will in general be much larger than the

10   magnitude of Q_{S2}.

The sum over N_1 values of I'_j in the subset $\{I'_j^{pos I}, Q'_j^{pos I}\}$ corresponding to the positive subset $\{I_j^{pos I}, Q_j^{pos I}\}$ is, from Eq. (1) above:

$$I'_{S1} = BI_{S1} - BTQ_{S1} + N_1K_i \tag{3},$$

and the sum over N_1 values of I'_j in the subset $\{I'_j^{neg I}, Q'_j^{neg I}\}$ corresponding to the

15   negative subset $\{I_j^{neg I}, Q_j^{neg I}\}$ is, also from Eq. (1):

$$I'_{S2} = BI_{S2} - BTQ_{S2} + N_1K_i \tag{4}.$$

Subtracting Eq. (4) from Eq. (3), K_i is eliminated, giving:

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$$B = \frac{(I'_{S1}-I'_{S2})+BT(Q_{S1}-Q_{S2})}{(I_{S1}-I_{S2})} = \frac{(I'_{S1}-I'_{S2})}{(I_{S1}-I_{S2})}\left[1-\frac{T(Q_{S1}-Q_{S2})}{(I_{S1}-I_{S2})}\right]^{-1}.$$

Because T is small and $(Q_{S1} - Q_{S2})$ is much smaller than $(I_{S1} - I_{S2})$, an adequate solution for B is obtained by substituting the old (or the initial) value of T in the right hand side of the above expression for B.

Next, the desired I_j and Q_j are divided into two subsets, the first subset $\{I_j^{posQ}, Q_j^{posQ}\}$ containing only the positive values of Q_j, and the second subset $\{I_j^{negQ}, Q_j^{negQ}\}$ containing only the negative values of Q_j. If one subset contains more values than the other subset, only N_2 values (where N_2 is equal to the number of values in the smaller subset) are used in both subsets.

The sum over N_2 values of I_j in the positive subset $\{I_j^{posQ}, Q_j^{posQ}\}$ is defined to be I_{S3}:

$$I_{S3} = \sum_{j=1}^{N_2} I_j^{posQ}$$

and the sum over N_2 values of Q_j in the positive subset $\{I_j^{posQ}, Q_j^{posQ}\}$ is defined to be Q_{S3}:

$$Q_{S3} = \sum_{j=1}^{N_2} Q_j^{posQ}.$$

The sum over N_2 values of I'_j in the subset $\{I'_j^{posQ}, Q'_j^{posQ}\}$ corresponding to the positive subset $\{I_j^{posQ}, Q_j^{posQ}\}$ is defined to be I'_{S3}:

$$I'_{S3} = \sum_{j=1}^{N_2} I'_j^{posQ}$$

and the sum over N_2 values of Q'_j in the subset $\{I'_j^{posQ}, Q'_j^{posQ}\}$ corresponding to the positive subset $\{I_j^{posQ}, Q_j^{posQ}\}$ is defined to be Q'_{S3}:

Because all of the Q values summed to get Q_{S3} were chosen to be positive and the signs of the corresponding I values summed to get I_{S3} were

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$$Q'_{S3} = \sum_{j=1}^{N_2} Q'^{posQ}_j.$$

uncorrelated, Q_{S3} will in general be much larger than I_{S3}.

The sum over N_2 values of I_j in the negative subset $\{I_j^{negQ}, Q_j^{negQ}\}$ is defined to be I_{S4}:

$$I_{S4} = \sum_{j=1}^{N_2} I_j^{negQ}$$

and the sum over N_2 values of Q_j in the negative subset $\{I_j^{negQ}, Q_j^{negQ}\}$ is defined to be Q_{S4}:

$$Q_{S4} = \sum_{j=1}^{N_2} Q_j^{negQ}.$$

In a similar way, the sum over N_2 values of I'_j in the subset $\{I'_j^{negQ}, Q'_j^{negQ}\}$ corresponding to the negative subset $\{I_j^{negQ}, Q_j^{negQ}\}$ is defined to be I'_{S4}:

$$I'_{S4} = \sum_{j=1}^{N_2} I'^{negQ}_j$$

and the sum over N_2 values of Q'_j in the subset $\{I'_j^{negQ}, Q'_j^{negQ}\}$ corresponding to the negative subset $\{I_j^{negQ}, Q_j^{negQ}\}$ is defined to be Q'_{S4}:

$$Q'_{S4} = \sum_{j=1}^{N_2} Q'^{negQ}_j.$$

Because all of the Q values summed to get Q_{S4} were chosen to be negative and the signs of the corresponding I values summed to get I_{S4} were uncorrelated, the magnitude of Q_{S4} will in general be much larger than the magnitude of I_{S4}.

The sum over N_2 values of Q'_j in the subset $\{I'^{posQ}_j, Q'^{posQ}_j\}$ corresponding to the positive subset $\{I^{posQ}_j, Q^{posQ}_j\}$ is, from Eq. (2):

$$Q'_{s3} = CQ_{s3} - CTI_{s3} + N_2K_q \qquad (5),$$

and the sum over N_2 values of Q'_j in the subset $\{I'^{negQ}_j, Q'^{negQ}_j\}$ corresponding to the negative subset $\{I^{negQ}_j, Q^{negQ}_j\}$ is, also from Eq. (2):

$$Q'_{s4} = CQ_{s4} - CTI_{s4} + N_2K_q \qquad (6).$$

Subtracting Eq. (6) from Eq. (5), K_q is eliminated, giving:

$$C = \frac{(Q'_{s3} - Q'_{s4}) + CT(I_{s3} - I_{s4})}{(Q_{s3} - Q_{s4})} = \frac{(Q'_{s3} - Q'_{s4})}{(Q_{s3} - Q_{s4})} \left[1 - \frac{T(I_{s3} - I_{s4})}{(Q_{s3} - Q_{s4})}\right]^{-1}.$$

Because T is small and $(I_{s3} - I_{s4})$ is much smaller than $(Q_{s3} - Q_{s4})$, an adequate solution for C is obtained by substituting the old (or the initial) value of T in the right hand side of the above expression for C.

Similarly, the sum over N_1 values of Q'_j in the subset $\{I'^{posI}_j, Q'^{posI}_j\}$ corresponding to the positive subset $\{I^{posI}_j, Q^{posI}_j\}$ is, from Eq. (2):

$$Q'_{s1} = CQ_{s1} - CTI_{s1} + N_1K_q \qquad (7),$$

and the sum over N_1 values of Q'_j in the subset $\{I'^{negI}_j, Q'^{negI}_j\}$ corresponding to the negative subset $\{I^{negI}_j, Q^{negI}_j\}$ is, also from Eq. (2):

$$Q'_{s2} = CQ_{s2} - CTI_{s2} + N_1K_q \qquad (8).$$

Subtracting Eq. (8) from Eq. (7), K_q is eliminated, giving:

$$T = \frac{(Q'_{s2} - Q'_{s1}) + C(Q_{s1} - Q_{s2})}{C(I_{s1} - I_{s2})}.$$

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Furthermore, the sum over N_2 values of I'_j in the subset $\{I'_j{}^{posQ}$, $Q'_j{}^{posQ}\}$ corresponding to the positive subset $\{I_j{}^{posQ}, Q_j{}^{posQ}\}$ is, from Eq. (1):

$$I'_{s3} = BI_{s3} - BTQ_{s3} + N_2K_i \qquad (9),$$

and the sum over N_2 values of I'_j in the subset $\{I'_j{}^{negQ}, Q'_j{}^{negQ}\}$ corresponding to the negative subset $\{I_j{}^{negQ}, Q_j{}^{negQ}\}$ is, also from Eq. (1):

$$I'_{s4} = BI_{s4} - BTQ_{s4} + N_2K_i \qquad (10).$$

Subtracting Eq. (10) from Eq. (9), K_i is eliminated, giving:

$$T = \frac{(I'_{s4}-I'_{s3})+B(I_{s3}-I_{s4})}{B(Q_{s3}-Q_{s4})}.$$

Using the updated solutions for B and C above in the expressions for T, an updated solution for T is obtained. Finally, substituting the new values A^{new}, B^{new}, and T^{new} in Eqs. (1) and (2), and summing over all N values if I_j, Q_j, I'_j, and Q'_j, new solutions are obtained for K_i:

$$K_i^{new} = \frac{1}{N}\sum_{j=1}^{N}\left[I'_j+B^{new}(T^{new}Q_j-I_j)\right],$$

and for K_q:

$$K_q^{new} = \frac{1}{N}\sum_{j=1}^{N}\left[Q'_j+C^{new}(T^{new}I_j-Q_j)\right].$$

SELF-ADJUSTING QUADRATURE MODULATORS

A self-adjusting quadrature modulator implementing the above-described procedure is shown in Fig. 3. A first digital signal processor 110 receives an information signal to be transmitted and converts the information to I and Q waveforms according to the intended modulation technique. The I and Q waveforms are converted from the numerical values produced by the digital signal

processor 110 to analog waveforms using digital-to-analog (D/A) converters 112,
113 (for the I and Q waveforms, respectively) as required by a quadrature
modulator 114. Aspects of quadrature modulators and the advantages of high-bit-
rate delta-sigma modulation are described in U.S. Patent Application No.
5   07/967,027 incorporated by reference above. High-bit-rate delta-sigma-modulation
bitstreams are easily converted to the analog voltage they represent by forming the
moving average voltage over a large number of bits. This can be done by a
continuous-time, low-pass filter having a bandwidth that is a small fraction of the
bit rate but that is still sufficient to pass all desired modulation components. For a
10   balanced signal configuration, balanced filters would be used.

Quadrature modulator integrated circuits are commercially available,
for example, from Hewlett-Packard Co. (part no. MX2001), and from Siemens
(part no. PMB2200). These circuits have balanced I and Q inputs. If, instead of
using high-bit-rate delta-sigma modulation to convert numerical I, Q values to
15   analog waveforms, a conventional D/A converter such as an 8- or 12-bit device
were employed, then either four matched devices would be needed to drive the $\pm$I
and $\pm$Q inputs of the modulator, or a pair of devices having balanced outputs.
However, use of the delta-sigma technique can be integrated as a small part of a
larger digital integrated circuit, and can avoid the complications associated with
20   the use of conventional D/A converters.

An upconverter 115, comprising a mixer and suitable bandpass
filters, translates the output of the quadrature modulator 114 from an intermediate
frequency, at which the quadrature modulator 114 most conveniently operates, to
the frequency of transmission. Power amplifiers 116, 117 raise the power level to
25   the desired transmission value. A coupler 118 extracts a sample of the modulated
transmission signal from any convenient point in the post-modulation transmission
chain. In Fig. 3, since the sample is extracted at the final frequency just before
the final transmission power amplifier 117, the sample is translated down to a
suitable frequency for comparison with the intended modulation by a
30   downconverter 119. A local oscillator frequency synthesizer 120 may
conveniently be used to drive both the upconverter 115 and the downconverter
119. Alternatively, downconversion can be effected by the downconverter 119

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and a different local oscillator frequency synthesizer provided the different
synthesizer synthesizes its frequency using the output of a reference frequency
standard 121 from which all other frequencies f_1 through f_8 that are used as shown
are derived. It will be understood that sampling the signal late in the post-

5   modulation transmission chain permits correction of errors arising in post-
modulator components.

The downconverted signal sample extracted from the post-modulated
transmission signal by the coupler 118 is subjected to a log-polar digitization
utilizing an intermediate frequency amplifier 122, which produces an output signal

10   approximately proportional to the natural logarithm of the instantaneous amplitude
of the signal sample as well as a hardlimited signal that preserves the instantaneous
signal phase information. The logamplitude signal is digitized by a suitable
analog-to-digital (A/D) converter 123 and the hardlimited, phase-preserving signal
is digitized by a suitable phase digitizer 124. The phase digitizer 124 can

15   advantageously be constructed as described in U.S. Patent No. 5,148,373 which is
expressly incorporated here by reference.

The A/D converter 123 may be of the successive approximation
type with an accuracy of eight bits. Alternatively, the converter 123 could first
employ high-bit-rate delta-sigma modulation to digitize the signal, followed by a

20   decimation filter to downsample the high-bit-rate delta-sigma bitstream to a lower-
rate stream of binary numbers. High bit-rate delta modulation or alternatively
companded delta modulation could also be used, having an implicit differentiation
(i.e., it measures rate of change of the log-amplitude signal) that has to be undone
by re-integrating numerically afterwards. The latter technique has the advantage

25   that small amplitude changes can more easily be resolved, a factor which could be
important for modulations that have or should have little amplitude modulation
component.

The log-polar digitized signal samples output from the A/D
converter 123 and the phase digitizer 124 are fed to a second digital signal

30   processor 125 that also receives from the first digital signal processor 110 the
desired (uncorrected) I and Q modulations. The second digital signal processor
125 performs phase alignment of the signal samples by modulo-2π adding a

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numerical phase offset value to the phase samples before log-polar-to-cartesian conversion. The second digital signal processor 125 then compares the phase-aligned, log-polar-to-cartesian-converted signal samples with the desired I and Q modulation values using a suitable process such as that described above to determine correction factors that are fed back to the first digital signal processor 110. The first processor 110 uses the correction factors to generate corrected, self-adjusted I and Q waveforms for modulation and transmission. The phase alignment constant ($\pm \delta\phi$) may also be updated by techniques similar to those already described above, and successive cycles will yield successive corrections. Of course, the functions of the processors 110, 125 could be carried out by a suitably capable signal processor. A suitable digital signal processing chip is, for example, the model no. TMS320C50, made by Texas Instruments, which can operate at instruction speeds of at least 20 MIPS.

It will be appreciated that the non-ideal features and imperfections of I and Q modulators are substantially constant for substantial periods of time, or change very slowly, for example, due to temperature changes, and thus the correction factors do not need to be continuously updated. On the other hand, the correction factors can be updated almost continuously if conditions require it, and even occasional updating of the correction factors will result in a system that is far more responsive, interactive, and accurate than previously known once-and-for-all, fixed compensating modulator adjustments. The present invention is particularly advantageous when used with a subtractive CDMA system using orthogonal coding, as described in the U.S. patent and patent application incorporated by reference above. CDMA demodulation techniques will now be described in conjunction with the signal graphs shown in Figs. 4-6, which set forth example waveforms in the coding and decoding processes involved in traditional CDMA systems. Using the waveform examples from Figs. 4-6, the improved performance of a subtractive CDMA demodulation technique is illustrated in Fig. 7.

Two different datastreams, shown in Fig. 4 as signal graphs (a) and (d), represent digitized information to be communicated over two separate communication channels. Information signal 1 is modulated using a high bit rate,

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digital code that is unique to signal 1 and that is shown in signal graph (b). For purposes of this description, the term "bit" refers to a binary digit or symbol of the information signal. The term "bit period" refers to the time period between the start and the finish of one bit of the information signal. The term "chip" refers to a binary digit of the high rate code signal. Accordingly, the term "chip period" refers to the time period between the start and the finish of one chip of the code signal. Naturally, the bit period is much greater than the chip period. The result of this modulation, which is essentially the product of the two signal waveforms, is shown in the signal graph (c). In Boolean notation, the modulation of two binary waveforms is essentially an exclusive-OR operation. A similar series of operations is carried out for information signal 2 as shown in signal graphs (d)-(f). In practice, of course, many more than two coded information signals are spread across the frequency spectrum available for cellular telephone communications.

Each coded signal is used to modulate an RF carrier using any one of a number of modulation techniques, such as QPSK. In a cellular telephone system, each modulated carrier is transmitted over an air interface. At a radio receiver, such as a cellular base station, all of the signals that overlap in the allocated frequency bandwidth are received together. The individually coded signals are added, as represented in the signal graphs (a)-(c) of Fig. 5, to form a composite signal waveform (graph (c)).

After demodulation of the received signal to the appropriate baseband frequency, the decoding of the composite signal takes place. Information signal 1 may be decoded or despread by multiplying the received composite signal shown in Fig. 5(c) with the unique code used originally to modulate signal 1 that is shown in signal graph (d). The resulting signal is analyzed to decide the polarity (high or low, +1 or -1, "1" or "0") of each information bit period of the signal. The details of how the receiver's code generator becomes time synchronized to the transmitted code are known in the art.

These decisions may be made by taking an average or majority vote of the chip polarities during each bit period. Such "hard" decision making processes are acceptable as long as there is no signal ambiguity. For example,

during the first bit period in the signal graph (f), the average chip value is +1.00 which readily indicates a bit polarity +1. Similarly, during the third bit period, the average chip value is +0.75, and the bit polarity is also most likely a +1. However, in the second bit period, the average chip value is zero, and the

5   majority vote or average test fails to provide an acceptable polarity value.

In such ambiguous situations, a "soft" decision making process must be used to determine the bit polarity. For example, an analog voltage proportional to the received signal after despreading may be integrated over the number of chip periods corresponding to a single information bit. The sign or polarity of the net

10   integration result indicates that the bit value is a +1 or -1.

The decoding of signal 2, similar to that of signal 1, is illustrated in the signal graphs (a)-(d) of Fig. 6. However, after decoding, there are no ambiguous bit polarity situations.

Theoretically, this decoding scheme can be used to decode every

15   signal that makes up the composite signal. Ideally, the contribution of unwanted interfering signals is minimized when the digital spreading codes are orthogonal to the unwanted signals. (Two binary sequences are orthogonal if they differ in exactly one half of their bit positions.) Unfortunately, only a certain number of orthogonal codes exist for a given word length. Another problem is that

20   orthogonality can be maintained only when the relative time alignment between two signals is strictly maintained. In communications environments where portable radio units are moving constantly, such as in cellular systems, precise time alignment is difficult to achieve. When code orthogonality cannot be guaranteed, noise-based signals may interfere with the actual bit sequences

25   produced by different code generators, e.g., the mobile telephones. In comparison with the originally coded signal energies, however, the energy of the noise signals is usually small.

"Processing gain" is a parameter of spread spectrum systems, and for a direct spreading system it is defined as the ratio of the spreading or coding

30   bit rate to the underlying information bit rate, i.e., the number of chips per information bit or symbol. Thus, the processing gain is essentially the bandwidth spreading ratio, i.e., the ratio of the bandwidths of the spreading code and

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information signal. The higher the code bit rate, the wider the information is
spread and the greater the spreading ratio. For example, a one kilobit per second
information rate used to modulate a one megabit per second code signal has
processing gain of 1000:1. The processing gain shown in Fig. 4, for example, is

5   8:1, the ratio of the code chip rate to the information datastream bit rate.

        Large processing gains reduce the chance of decoding noise signals
modulated using uncorrelated codes. For example, processing gain is used in
military contexts to measure the suppression of hostile jamming signals. In other
environments, such as cellular systems, processing gain helps suppress other,

10   friendly signals that are present on the same communications channel but use
codes that are uncorrelated with the desired code. In the context of the subtractive
CDMA demodulation technique, "noise" includes both hostile and friendly signals,
and may be defined as any signals other than the signal of interest, i.e., the signal
to be decoded. Expanding the example described above, if a signal-to-interference

15   ratio of 10:1 is required and the processing gain is 1000:1, conventional CDMA
systems have the capacity to allow up to 101 signals of equal energy to share the
same channel. During decoding, 100 of the 101 signals are suppressed to
1/1000th of their original interfering power. The total interference energy is thus
100/1000, or 1/10, as compared to the desired information energy of unity. With

20   the information signal energy ten times greater than the interference energy, the
information signal may be correlated accurately.

        Together with the required signal-to-interference ratio, the
processing gain determines the number of allowed overlapping signals in the same
channel. That this is still the conventional view of the capacity limits of CDMA

25   systems may be recognized by reading, for example, the above-cited paper by
Gilhousen et al.

        In contrast to conventional CDMA, an important aspect of the
subtractive CDMA demodulation technique is the recognition that the suppression
of friendly CDMA signals is not limited by the processing gain of the spread

30   spectrum demodulator as is the case with the suppression of military type jamming
signals. A large percentage of the other signals included in a received, composite
signal are not unknown jamming signals or environmental noise that cannot be

25

correlated. Instead, most of the noise, as defined above, is known and is used to facilitate decoding the signal of interest. The fact that the characteristics of most of these noise signals are known, including their corresponding spreading codes, is used in the subtractive CDMA demodulation technique to improve system capacity and the accuracy of the signal decoding process. Rather than simply decode each information signal from the composite signal, the subtractive CDMA demodulation technique also removes each information signal from the composite signal after it has been decoded. Those signals that remain are decoded only from the residual of the composite signal. Consequently, the already decoded signals do not interfere with the decoding of the remaining signals.

For example, in Fig. 7, if signal 2 has already been decoded as shown in the signal graph (a), the coded form of signal 2 can be reconstructed as shown in the signal graphs (b) and (c) (with the start of the first bit period of the reconstructed datastream for signal 2 aligned with the start of the fourth chip of the code for signal 2 as shown in Fig. 4 signal graphs (d) and (e)), and subtracted from the composite signal in the signal graph (d) (again with the first chip of the reconstructed coded signal 2 aligned with the fourth chip of the received composite signal) to leave coded signal 1 in the signal graph (e). This is easily verified by comparing signal graph (e) in Fig. 7 with signal graph (c) in Fig. 4 (truncated by removing the first three and the very last chip). Signal 1 is recaptured easily by multiplying the coded signal 1 with code 1 to reconstruct signal 1. Note that because the bit periods of datastreams for signals 1 and 2 are shifted relative to one another by 2 chips there are only six +1 chips in the first bit period of the recaptured signal 1 shown in Fig. 7 signal graph (f). It is significant that while the conventional CDMA decoding method was unable to determine whether the polarity of the information bit in the second bit period of signal 1 was a +1 or a -1 in signal graph (f) of Fig. 5, the decoding method of the subtractive CDMA demodulation technique effectively resolves that ambiguity simply by removing signal 2 from the composite signal.

A general CDMA system will now be described in conjunction with Figs. 8(a), 8(b). In the transmitter shown in Fig. 8(a), an information source such as speech is converted from analog format to digital format in a conventional

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source coder 20. The digital bitstream generated by the transmitter source coder
20 may be further processed in a transmitter error correction coder 22 that adds
redundancy which increases the bandwidth or bit rate of the transmission. In
response to a spreading code selection signal from a suitable control mechanism

5   such as a programmable microprocessor (not shown), a particular spreading code
is generated by a transmit spreading code generator 24, which may be a
pseudorandom number generator. The selected spreading code is summed in a
modulo-2 adder 26 with the coded information signal from the error correction
coder 22. It will be appreciated that the modulo-2 addition of two binary

10   sequences is essentially an exclusive-OR operation in binary logic. The modulo-2
summation effectively "spreads" each bit of information from the coder 22 into a
plurality of "chips".

            The coded signal output by the adder 26 is used to modulate an RF
carrier using a modulation technique such as QPSK in a modulator 28. The

15   modulated carrier is transmitted over an air interface by way of a conventional
radio transmitter 30. A plurality of the coded signals overlapping in the allocated
frequency band are received together in the form of a composite signal waveform
at a radio receiver 32, such as a cellular radiotelephone base station, illustrated in
Fig. 8(b). After demodulation in a demodulator 34 to baseband, the composite

20   signal is decoded.

            An individual information signal is decoded or "despread" by
multiplying the composite signal with the corresponding unique spreading code
produced by a receiver spreading code generator 36. This unique code
corresponds to that spreading code used originally to spread that information signal

25   in the transmit spreading code generator 24. The spreading code and the
demodulated signal are combined by a multiplier 38. Because several received
chips represent a single bit of transmitted information, the output signal of
multiplier 38 may be successively integrated over a particular number of chips in
order to obtain the actual values of the information bits. As described above,

30   these bit value decisions may be made by taking an average or majority vote of the
chip polarities during each bit period. In any event, the output signals of
multiplier 38 are eventually applied to a receiver error correction decoder 40 that

reverses the process applied by the transmitter error correction coder 22, and the resulting digital information is converted into analog format (e.g., speech) by a source decoder 42.

As described above, this decoding scheme theoretically can be used to decode every signal in the composite signal. Ideally, the contribution of unwanted, interfering signals is minimized when the digital spreading codes are orthogonal to these unwanted signals and when the relative timing between the signals is strictly maintained. Unfortunately, only a certain number of orthogonal codes exist for a finite word length, and in communications environments where portable radio units are moving constantly, such as in cellular systems, time alignment is difficult to achieve.

In a preferred CDMA communication system, the error correction coding is based on orthogonal or bi-orthogonal block coding of the information to be transmitted. In orthogonal block coding, a number of bits M to be transmitted are converted to one of 2^M 2^M-bit orthogonal codewords. Decoding an orthogonal codeword involves correlation with all members of the set of $N = 2^M$ codewords. The binary index of the codeword giving the highest correlation yields the desired information. For example, if a correlation of sixteen 16-bit codewords numbered 0-15 produces the highest correlation on the tenth 16-bit codeword, the underlying information signal is the 4-bit binary codeword 1010 (which is the integer 10 in decimal notation, hence, the index of 10). Such a code is also termed a [16,4] orthogonal block code and has a spreading ratio $R = 16/4 = 4$. By inverting all of the bits of the codewords, one further bit of information may be conveyed per codeword. This type of coding is known as bi-orthogonal block coding.

A significant feature of such coding is that simultaneous correlation with all the orthogonal block codewords in a set may be performed efficiently by means of a Fast Walsh Transform (FWT) device. In the case of a [128,7] block code, for example, 128 input signal samples are transformed into a 128-point Walsh spectrum in which each point in the spectrum represents the value of the correlation of the input signal samples with one of the codewords in the set. A programmable digital signal processor can readily be configured to calculate Walsh transforms, although use of the FWT is usually more efficient. A suitable FWT

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processor is described in commonly assigned U.S. Patent ~~Application No.~~ No. 5,357,454

~~07/735,805~~ filed on July 25, 1991, which is expressly incorporated here by

reference.

In a preferred embodiment of the present invention, communication

5   signals are first encoded into 7-bit bytes which are then further encoded using a

[128,7] orthogonal Walsh-Hadamard block code in order to produce 128-bit

codewords. The codewords for each particular signal are scrambled by modulo-2

addition of a scrambling mask that is unique to each signal. The scrambled

codewords are then transmitted bit serially by means of filtering and modulation.

10   A preferred system is described in commonly assigned U.S. Patent ~~Application~~ No. 07,866

~~No. 07/866,865~~ filed on April 10, 1992, entitled "Multiple Access Coding for

Mobile Radio Communications," which is expressly incorporated here by

reference.

The filtering and conversion into I and Q modulation waveforms is

15   preferably performed in the first digital signal processor 110 for all signals using

the same frequency channel. The I and Q waveforms are then added together with

a weighting factor depending on the relative signal strength with which each signal

is to be transmitted because it is logical and advantageous to transmit with a higher

signal strength to mobile stations farther away, at extreme range, while

20   transmitting at lower signal strength to mobile stations nearer the base station.

The summed I and Q waveforms are then subjected to the correction factors

described above before being output by the first digital signal processor to D/A

converters 112, 113 for subsequent quadrature modulation by the quadrature

modulator 114. The uncorrected I and Q values are also output from the first

25   digital signal processor 110 to the second digital signal processor 125 for

comparison with the measured values determined by the modulation assessment

receiver.

In this preferred implementation, receivers for the composite

CDMA signal radiated by a base station employ the subtractive CDMA technique

30   described above and in the above-cited U.S. patent and patent application. Each

mobile station decodes the strongest of the orthogonally coded signals first by

descrambling with the scrambling mask of the strongest signal, performing a 128-

29

point FWT, and determining the largest of the 128 transform components to detect which codeword was most likely to have been transmitted. The detected codeword is then subtracted from the composite signal, for example, by setting the largest transform component equal to zero, a 128-point Inverse Fast Walsh Transform

5   (IFWT) is performed, and finally the scrambling code is re-applied. The process is repeated successively on the residual composite signal using the descrambling code corresponding to the next strongest signal and so on until the mobile station has decoded the signal intended for it. In this way, the stronger signals are prevented from hindering the decoding of the weaker signals they overlap.

10   When a signal is subtracted, a waveform theoretically corresponding to a perfect modulation is subtracted, leaving as a residual any imperfections in the modulation process. According to this preferred embodiment of the present invention, a subtractive CDMA receiver is used as the modulation assessment receiver for correcting the transmitter modulation imperfections. Moreover, the

15   correction factors can be directly identified with certain transform components produced by the FWT.

The modulation employed to bit-serially transmit the 128-bit scrambled Walsh-Hadamard codewords is preferably Shaped Offset Quadrature Amplitude Modulation (SOQAM), which is related to OQPSK in that even bits are

20   applied to the I phase and odd bits are applied to the Q phase alternately. The sampling points for SOQAM are shown in the I and Q waveforms illustrated in Fig. 9. The desired sampling points shown in Fig. 9 are used in a conventional way to determine the characteristics of the clock signals f_s, f_t produced by the reference frequency and timing generator 121. OQPSK is further described in the

25   publication by S. Gronemeyer et al. cited above.

At the receiver, signal samples must, in principle, be taken from the I and Q channels alternately to obtain 128-sample blocks (representing one of the 128 128-bit Walsh-Hadamard block codewords) upon which the FWT is performed. Nevertheless, by applying a pre-rotation of the phase of the received

30   signal amounting to a progressive 90° per bit period, the phases of the even bits are rotated by 0° or 180°, which leaves the even bits in the I channel albeit with half of the even bits inverted, and the phases of the odd bits are rotated by 90° or

270°, which rotates the odd bits from the Q channel into the I channel. After such a pre-rotation, all 128 samples for the FWT can be collected from the same channel (in this example, the I channel). The effect of the pre-rotation is to change the signs of the samples according to the pattern

5      $+ + - - + + - - + + - -$ . . ., which is the Walsh-Hadamard basis codeword W2. Accordingly, the detected codeword will be offset by a bitwise modulo-2 addition of two from the codeword transmitted by virtue of the mathematical properties of Walsh-Hadamard codewords. It is a simple matter to correct the detected codeword by bitwise modulo-2 subtraction of the offset two. For example, if the

10   decimal data block 73 (binary number 01001001) is transmitted by sending the 128-bit Walsh-Hadamard codeword W73, the data block 75 (binary number 01001011) will be detected due to the pre-rotation effect. Bitwise modulo-2 addition of binary number 00000010 corrects the inverted data bit.

       Carrier leakage or I and Q offsets in the quadrature modulator

15   would appear in the transmitted signal as a constant carrier component, which corresponds to unscrambled Walsh-Hadamard codeword W0, but due to pre-rotation this carrier leakage component is converted to correspond to unscrambled Walsh-Hadamard codeword W2. Thus, such leakage or offsets can be detected by descrambling the composite signal with a null scrambling code, performing an

20   FWT, and identifying transform component 2 as the modulator imbalance. This transform component, which may be complex, contains the arbitrary phase introduced by the transmission path, which can be removed by relating it to the known phase of one of the transmitted signals.

       In a preferred system, the strongest of the overlapping signals is

25   used as a broadcast (calling) channel communicating with all the mobile stations, and the strongest of the overlapping signals is also used as a pilot or phase-reference signal to which the phase of the other signals and the above-mentioned imbalance measurement can be related. Specifically, if the complex value of the largest detected transform component when decoding the strongest signal is S_1, and

30   the measurement of the W2 transform component representing the modulator imbalance yields the complex number K, then the I and Q leakage components K_i

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and K_q, respectively, returned as correction factors to the first digital signal processor 110 are given by:

$$K_i + iK_q = S_1^* K = (Re(S_1) - iIm(S_1))(Re(K) + iIm(K)) .$$

For the case of relative scaling errors in the I and Q channels, if the I channel modulator produces a larger signal than the Q channel modulator, then alternate signal samples originally in the Q channel will be of smaller amplitude than the intervening signal samples coming from the I channel. This is equivalent to the addition of a small amount of a spurious codeword that differs from the correct codeword by having its Q bits inverted, thereby producing a small degree of destructive addition with the Q bits and a small degree of constructive addition with the I bits. This spurious codeword is thus the transmitted codeword modified by the sign pattern $+ - + - + - + - + - \ldots$, which is the Walsh-Hadamard basis codeword W1. Accordingly, the spurious codeword will be offset by a bitwise modulo-2 addition of one from the codeword transmitted, again by virtue of the mathematical properties of Walsh-Hadamard codewords. Accordingly, by determining the component of the Walsh-Hadamard transform one away from the transmitted codeword, a mis-scaling between I and Q channels can be identified and corrected.

It will be appreciated that this small spurious component may well be masked by spurious components arising from other signals, but if these are decoded and subtracted first the small error component can more easily be detected. Moreover, separate assessments of the value of the component representing I and Q relative mis-scaling can be made relative to the main decoded codeword after decoding of each of the overlapping signals. The values can then be averaged over all the codewords decoded from one block of 128 signal samples, as well as averaged over many signal blocks in order to average out the above-mentioned sources of spurious error that would otherwise mask the small component representing I and Q relative mis-scaling.

The component of the average that is in-phase with the transmitted codeword that is detected represents the mis-scaling factor A, while the component in quadrature to detected codewords represents the correction factor T for errors in

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the 90°-phase-splitting network 103 in the quadrature modulator. Thus, by intelligently using information provided by the Walsh transform signal decoder, all the errors in the quadrature modulator can be determined and appropriate correction factors transmitted to the modulating first digital signal processor 110 in
5   order for the first digital signal processor to take the corrective actions described above. In this way, it is possible to ensure that the composite modulation produced by the base station accurately conforms to the desired theoretical form.

While particular embodiments of the present invention have been described and illustrated, it should be understood that the invention is limited only
10   by the attached claims. Since modifications may be made by persons skilled in the art, the present application contemplates any and all modifications that fall within the spirit and scope of the invention disclosed and claimed herein.

THE CLAIMS DEFINING THE INVENTION ARE AS FOLLOWS:

1.   A self-adjusting modulator for providing an accurately modulated signal, comprising:

first digital signal processing means for numerically generating in-phase and quadrature modulation waveform samples and applying correction factors to numerically generated in-phase and quadrature modulation waveform samples;

a quadrature modulator for impressing an in-phase modulation waveform, corresponding to the numerically generated in-phase modulation waveform samples, on a substantially cosinusoidal carrier wave and for impressing a quadrature modulation waveform, corresponding to the numerically generated quadrature modulation waveform samples, on a substantially sinusoidal carrier wave;

transmitter means for upconverting and amplifying an output of said quadrature modulator to a predetermined transmission frequency and power level;

modulation assessment receiving means, coupled to said transmitter means, for collecting and digitising samples of the upconverted and amplified output of said quadrature modulator; and

second digital signal processing means, in communication with said first digital signal processing means and with said modulation assessment receiving means, for determining the correction factors based on said digitised samples from said modulation assessment receiving means and on uncorrected in-phase and quadrature modulation waveform samples from said first digital signal processing means, and for communicating the correction factors to the first digital signal processing means.

2.   The self-adjusting modulator of claim 1, further comprising at least one digital-to-analog converter for converting said numerically generated in-phase and quadrature waveform samples to said in-phase and quadrature modulation waveforms impressed by said quadrature modulator on said carrier waves.

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3. The self-adjusting modulator of claim 2, wherein said digital-to-analog converter uses substantially highly oversampled delta-sigma modulation.

4. The self-adjusting modulator of claim 1, wherein said digitized samples provided by said modulation assessment receiving means comprise samples in, alternatively, polar and log-polar formats.

5. The self-adjusting modulator of claim 4, wherein said second digital signal processing means converts said digitized samples to cartesian format.

6. The self-adjusting modulator of claim 1, wherein said in-phase and quadrature modulation waveforms comprise a spread-spectrum coded signal.

7. The self-adjusting modulator of claim 6, wherein said spread-spectrum coded signal comprises a sum of a plurality of spread-spectrum coded signals.

8. The self-adjusting modulator of claim 7, wherein said second digital signal processing means includes means for performing a Walsh transform in determining said correction factors.

9. The self-adjusting modulator of claim 7, wherein said spread-spectrum coded signals include a plurality of [128,7] Walsh-Hadamard codewords.

10. The self-adjusting modulator of claim 7, wherein said spread-spectrum coded signals include a plurality of [128,8] Walsh-Hadamard codewords.

11. The self-adjusting modulator of claim 1, wherein said second digital signal processing means includes means for performing a Walsh transform in determining said correction factors.

12.   The self-adjusting modulator of claim 1, wherein said accurately modulated signal comprises one of a Quadrature Phase Shift Keying signal, an Offset Quadrature Phase Shift Keying signal, a Shaped Offset Quadrature Phase Shift Keying signal, a Quadrature Amplitude Modulation signal, an Offset Quadrature Amplitude Modulation signal, and a Shaped Offset Quadrature Amplitude Modulation signal.

13.   A self-adjusting modulation method for providing an accurately modulated signal, comprising the steps of:

a)   numerically generating uncorrected in-phase and quadrature modulation waveform samples;

b)   applying a set of predetermined correction factors to said uncorrected in-phase and quadrature modulation waveform samples to produce corrected in-phase and quadrature modulation samples;

c)   forming a corrected in-phase modulation waveform corresponding to the corrected in-phase modulation samples and modulating a substantially cosinusoidal carrier wave with the corrected in-phase modulation waveform, forming a corrected quadrature modulation waveform corresponding to the corrected quadrature modulation samples and modulating a substantially sinusoidal carrier wave with the corrected quadrature modulation waveform, and forming a modulated output signal by summing the modulated cosinusoidal and sinusoidal carrier waves;

d)   upconverting and amplifying the modulated output signal to a selected desired transmission frequency and power level;

e)   collecting digital samples of the upconverted, amplified modulated output signal;

f)   processing said digital samples with the uncorrected in-phase and quadrature waveform samples to produce an updated set of correction factors; and

g)   successively repeating the foregoing steps a), b), c), and d).

14.   The method of claim 13, further comprising the step of converting said numerically generated in-phase and quadrature waveform samples to an analog form prior to step c).

15.   The method of claim 14, wherein said converting step comprises substantially highly oversampled delta-sigma modulation.

16.   The method of claim 13, wherein said digitised samples comprise, alternatively, polar and log-polar representations.

17.   The method of claim 16, wherein step d) includes the step of converting the, alternatively, polar and log-polar representations to cartesian representation.

18.   The method of claim 13, wherein said accurately modulated signal comprises a spread-spectrum coded signal.

19.   The method of claim 18, wherein step f) comprises the step of performing a Walsh transform on said digitised samples to determine said correction factors.

20.   The method of claim 18, wherein said spread-spectrum coded signal includes a plurality of [128,7] Walsh-Hadamard codewords.

21.   The method of claim 20, wherein said spread-spectrum signal includes a plurality of [128,8] Walsh-Hadamard codewords.

22.   The method of claim 13, wherein said accurately modulated signal comprises one of a Quadrature Phase Shift Keying signal, an Offset Quadrature Phase Shift Keying signal, a Shaped Offset Quadrature Phase Shift Keying signal, a Quadrature Amplitude Modulation signal, an Offset Quadrature

Amplitude Modulation signal, and a Shaped Offset Quadrature Amplitude Modulation signal.

23. The method of claim 13, wherein step f) includes the step of performing a Walsh transform on said digitised samples to determine said correction factors.

DATED this 25th day of June 1997.

ERICSSON INC.

WATERMARK PATENT & TRADEMARK ATTORNEYS
LEVEL 4, AMORY GARDENS
2 CAVILL AVENUE
ASHFIELD N.S.W. 2131
AUSTRALIA

LJD:MP:RM   DOC014   AU7056094.WPC
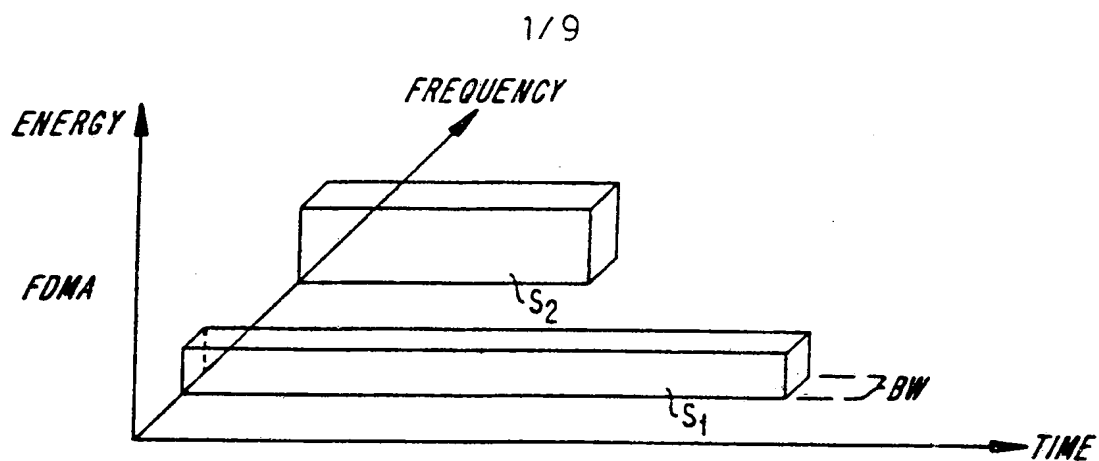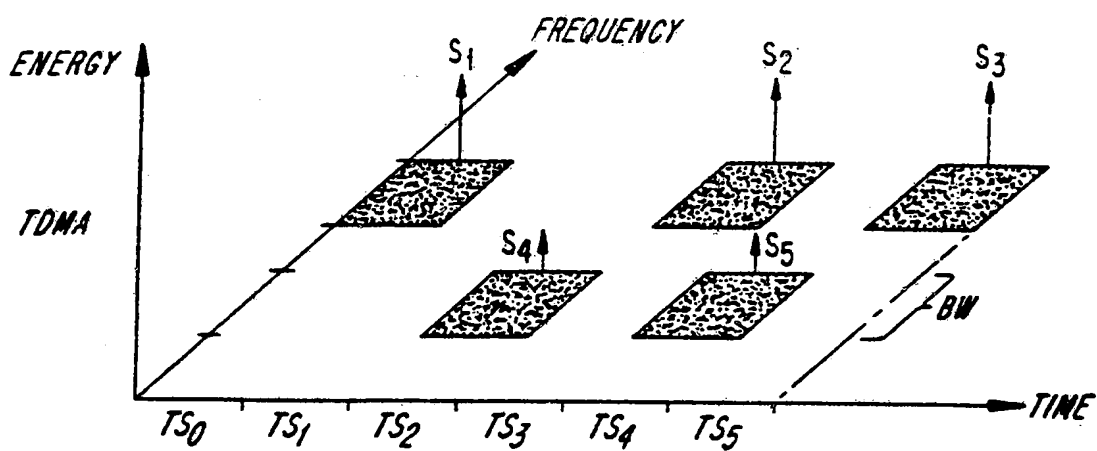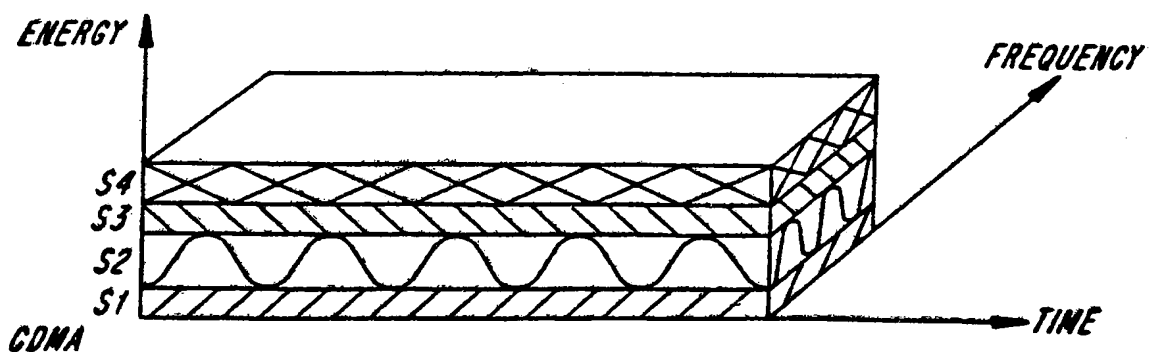
FIG. I a

FIG. I b

FIG. I c

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FIG. 2
PRIOR ART

OUTPUT

70560/94

FIG. 3
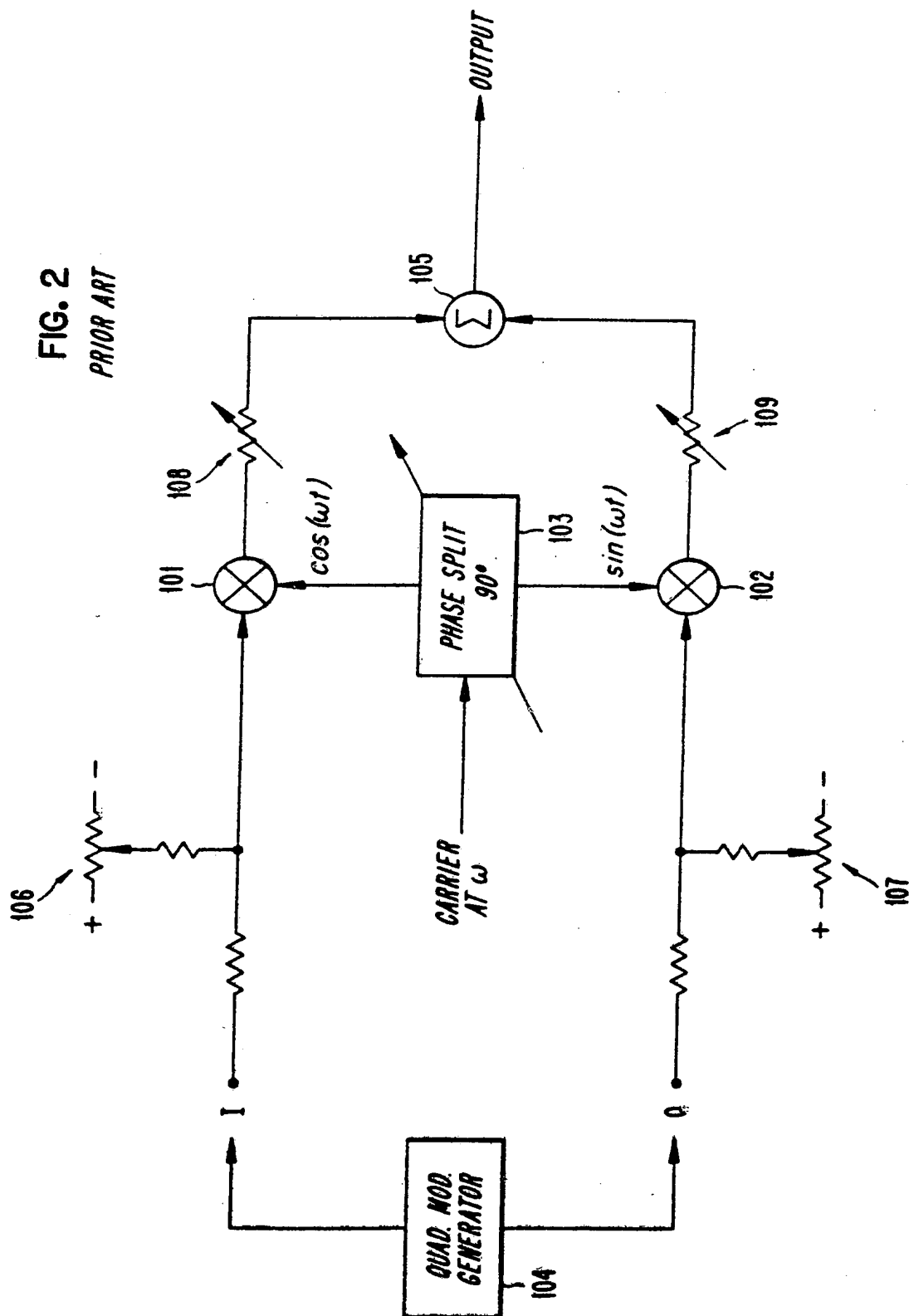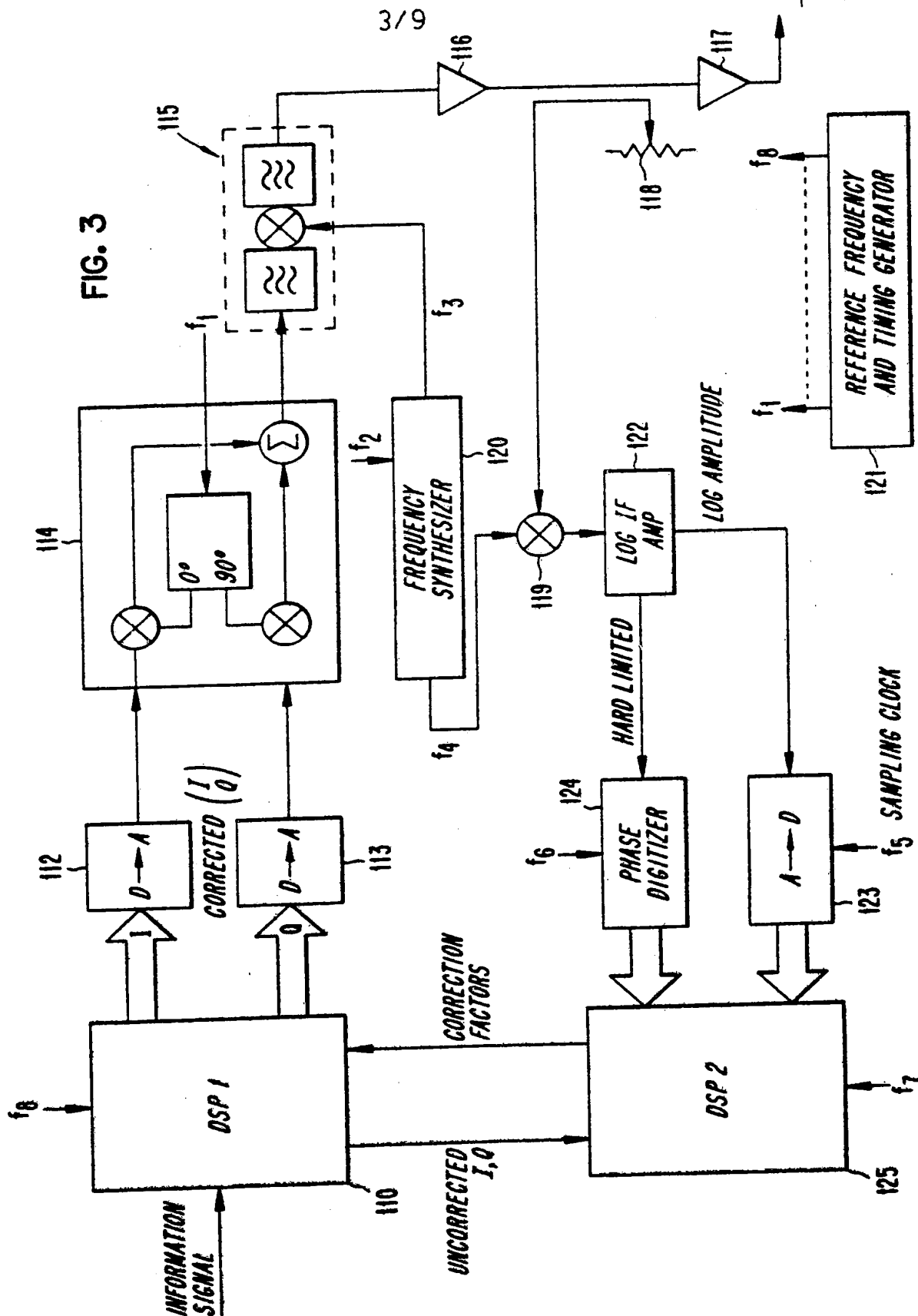
FIG. 4a DATASTREAM FOR SIGNAL 1.
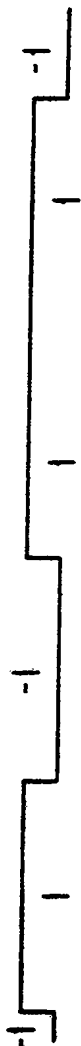
FIG. 4b CODE FOR SIGNAL 1.
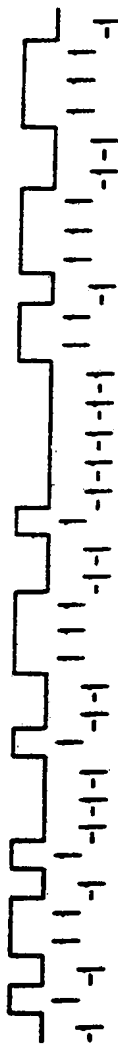
FIG. 4c PRODUCT OF SIGNAL 1 AND CODE 1.
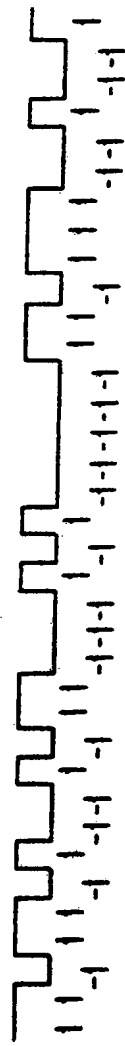
FIG. 4d DATASTREAM FOR SIGNAL 2.

FIG. 4e CODE FOR SIGNAL 2.
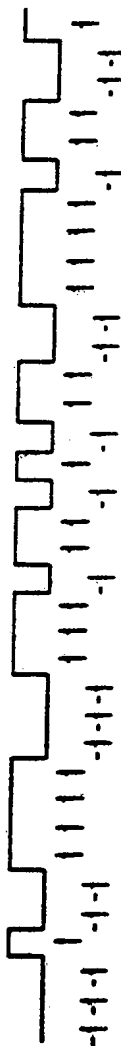
FIG. 4f PRODUCT OF SIGNAL 2 AND CODE 2.

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FIG. 5a   PRODUCT OF SIGNAL 1 AND CODE 1.

FIG. 5b   PRODUCT OF SIGNAL 2 AND CODE 2.

FIG. 5c   SUM OF CODED SIGNALS 1 AND 2.

FIG. 5d   CODE FOR SIGNAL 1.

FIG. 5e   PRODUCT OF SUM WITH SIGNAL 1'S CODE.

FIG. 5f
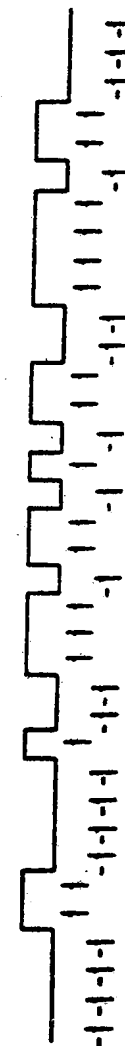
FIG. 6a   *SUM OF CODED SIGNALS 1 AND 2.*
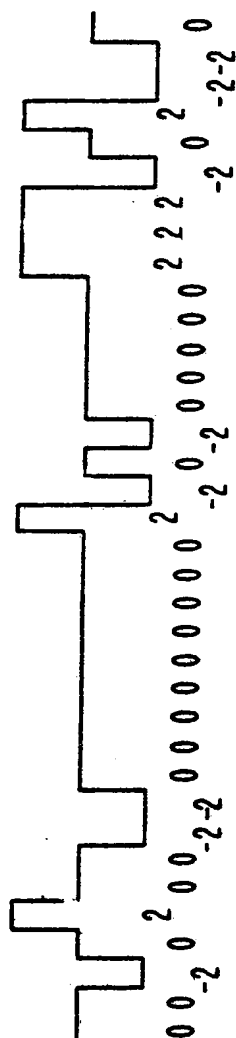
FIG. 6b   *CODE FOR SIGNAL 2.*

FIG. 6c   *PRODUCT OF SUM WITH SIGNAL 2'S CODE.*

FIG. 6d

AVERAGE = -0.75   AVERAGE = 0.25   AVERAGE = 0.5   AVERAGE = 1.75
DECISION = -   DECISION = +   DECISION = +   DECISION = +
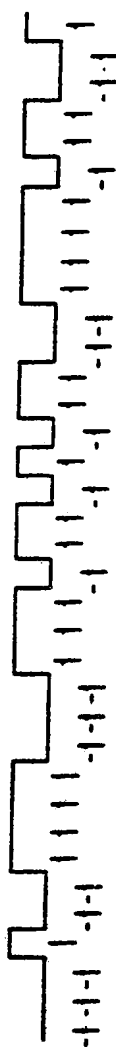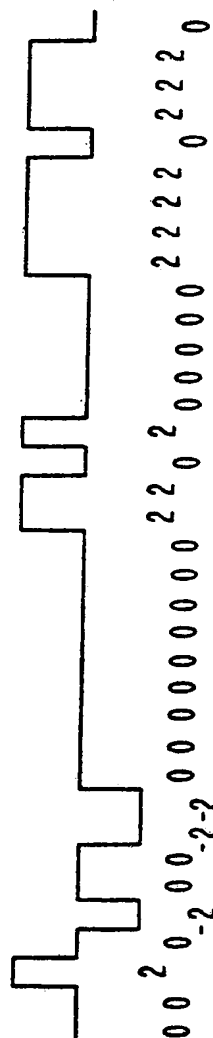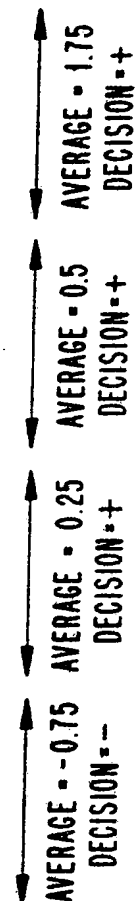
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FIG. 7a *RECONSTRUCTED DATASTREAM FOR SIGNAL 2.*

FIG. 7b *CODE FOR SIGNAL 2.*

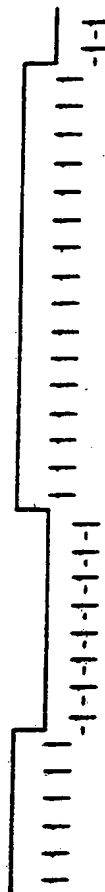**FIG. 7c** *RECONSTRUCTED CODED SIGNAL 2.*
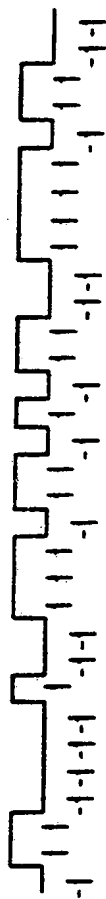
FIG. 7d *RECEIVED COMPOSITE SIGNAL.*
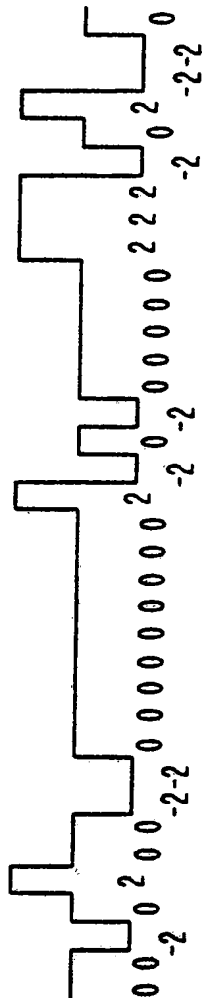
FIG. 7e *RESIDUAL OF SUM SIGNAL MINUS RECONSTRUCTED CODED SIGNAL 2.*

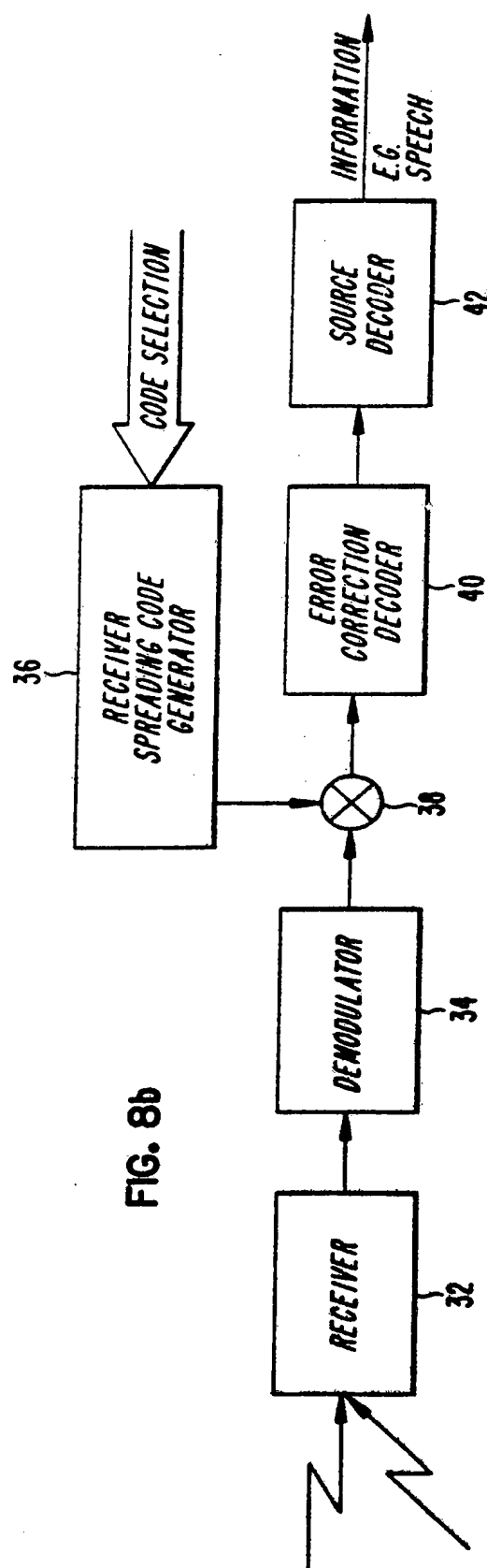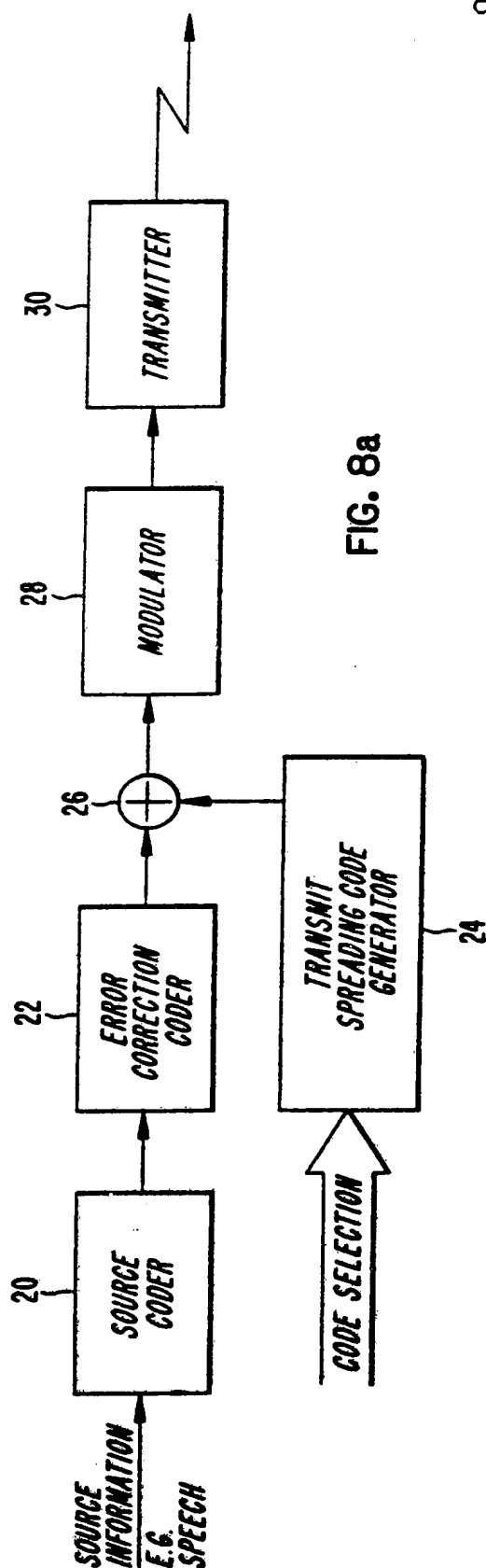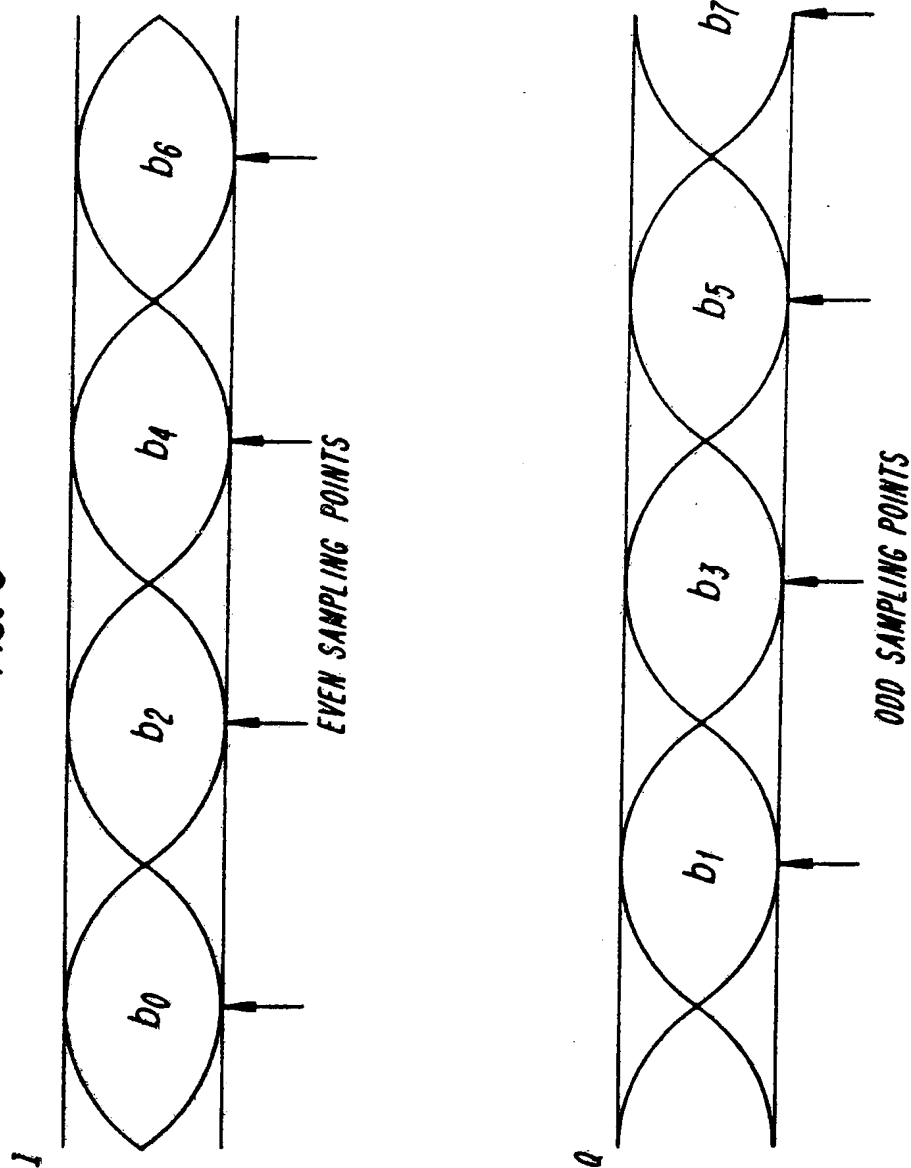**FIG. 7f** *PRODUCT OF SUBTRACTIVE RESIDUAL AND SIGNAL 1'S CODE.*

FIG. 8a

FIG. 8b

FIG. 9

EVEN SAMPLING POINTS

ODD SAMPLING POINTS

INTERNATIONAL SEARCH REPORT

A.	CLASSIFICATION OF SUBJECT MATTER
IPC(5)	Please See Extra Sheet
US CL	Please See Extra Sheet

According to International Patent Classification (IPC) or to both national classification and IPC

B.	FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U S   332/100, 101, 102, 103, 104, 105, 375/39, 60, 455/126

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C.	DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US, A, 5,012,208 (MAKINEN ET AL) 30 APRIL 1991, SEE THE ABSTRACT.	NONE

☐ Further documents are listed in the continuation of Box C.   ☐   See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance		
"E"	earlier document published on or after the international filing date	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"L"	document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)		
		"Y"	document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"O"	document referring to an oral disclosure, use, exhibition or other means		
"P"	document published prior to the international filing date but later than the priority date claimed	"&"	document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
10 AUGUST 1994	17 AUG 1994

Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	Authorized officer DAVID MIS
Facsimile No. N/A	Telephone No. (703) 308-4907

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A CLASSIFICATION OF SUBJECT MATTER
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H03C 3/00, 3/08

A CLASSIFICATION OF SUBJECT MATTER
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332/103; 375/60